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TE-MODE SOLUTIONS FOR DIELECTRIC SLAB CENTER LOADED RIDGED WAVE--ETC(U)
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NRL Report 8105

TE– Mode Solutions for Dielectric Slab Center Loaded Ridged Waveguide

CHARLES W. YOUNG, JR.

Microwave Techniques Branch
Electronics Technology Division

April 29, 1977

NAVAL RESEARCH LABORATORY
Washington, D.C.

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A Dielectric slab center loaded ridged waveguide has been analyzed to obtain TE_{10} propagation characteristics for the lossless case. The analysis was performed by solving for the ~~lowest~~ order resonance of the transverse wave component, using an equivalent transmission line circuit which models the height change at the ridge as an effective shunt susceptance. Calculated and measured data are compared; they show good agreement. The analysis should prove useful in designing transformers to match empty ridged waveguide to slab loaded rectangular waveguide components such as toroidal phase shifters.

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CONTENTS

TE-MODE SOLUTIONS FOR DIELECTRIC-SLAB CENTER-LOADED RIDGED WAVEGUIDE

INTRODUCTION

Applications exist which require dielectric or ferrite slab center loaded rectangular waveguide to be used in conjuction with ridged waveguide. Transitions can be made with stepped matching transformers; these transformers are appropriate sections of a composite of the two different waveguide configurations. The objective of this report is to present an analysis of such dielectric slab center loaded ridged waveguide and to provide a method of obtaining equations for the TE_{n0} propagation characteristics, thus facilitating transformer design.

BACKGROUND

Because higher order modes can easily cause mismatch and transmission loss spikes, waveguide operation is generally limited to a frequency band where only the principal mode may propagate. Conventional rectangular waveguide has a theoretical two-to-one, or single octave, principal-mode-only frequency bandwidth; in practice, the useable bandwidth is less because of large attenuation near the cut-off frequency.

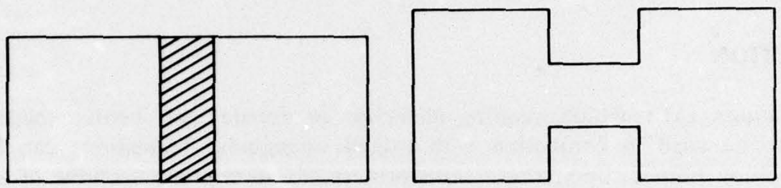
Ridged waveguide, particularly double-ridged waveguide, is commonly used when larger bandwidths are required at high power levels. A frequency range of more than four to one between the cut-off frequencies of the TE_{10} and TE_{20} modes can easily be obtained [1,2] using double-ridged waveguide. Similar bandwidths can be obtained with dielectric slab center loaded rectangular waveguide [3,4,5]. Both ridged and slab loaded waveguide achieve broad bandwidths by adding large capacitance to the dominant mode while only slightly affecting the capacitance of the next higher order mode.

Ferrite toroidal phase shifters also can be designed for operation in excess of one octave. Because of the small gap spacing of ridged waveguide, the phase shifters are generally made in rectangular waveguide. Dielectric slab center loaded rectangular waveguide would be readily compatible with the ferrite toroidal phase shifter, but it is not a commonly used transmission line. Since ridged waveguide is commonly used, it would be desirable to have compatibility, i.e., matching transitions, between ridged waveguide and ferrite toroidal loaded rectangular waveguide. Dielectric loaded tapered transitions are possible, but the fabrication would be very difficult. Also, a quasi-Tchebycheff transformer design should give better matching for given length transitions. The latter approach requires transformer sections of dielectric loaded ridged waveguide, but analysis of this type of transmission line is not currently available in the literature. The analysis in this report employs an equivalent transmission line circuit for the transverse component of the propagating electromagnetic wave to derive solutions for the TE_{n0} propagation constants in such waveguide.

Manuscript submitted January 27, 1977.

ANALYSIS

TE-mode solutions for the dielectric slab center loaded rectangular waveguide of Fig. 1a can be derived by using ABCD matrices [6] or by using an equivalent transmission line circuit for the crossguide component of the electromagnetic wave. The homogeneous double-ridged waveguide of Fig. 1b has been analyzed [1,2] by using the latter method in conjunction with the equivalent discontinuity susceptance due to the height change at the ridge wall.



(a) Dielectric slab center loaded rectangular waveguide

(b) Double-ridged waveguide

Fig. 1 — Broadband waveguide cross sections

For the dielectric slab center loaded double ridged waveguide of Fig. 2, the analysis is similar to that for the homogeneous case, with an extra section incorporated in the equivalent transmission line circuit. The dimensions referred to in all subsequent calculations are those shown in Fig. 2. For simplicity, this report will consider only the case for TE_{n0} modes and will assume that the transmission line is lossless, i.e., perfectly conducting waveguide walls and a *dielectric loss tangent of zero.* Axial symmetry will also be assumed.

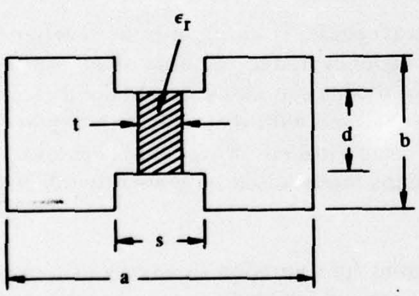
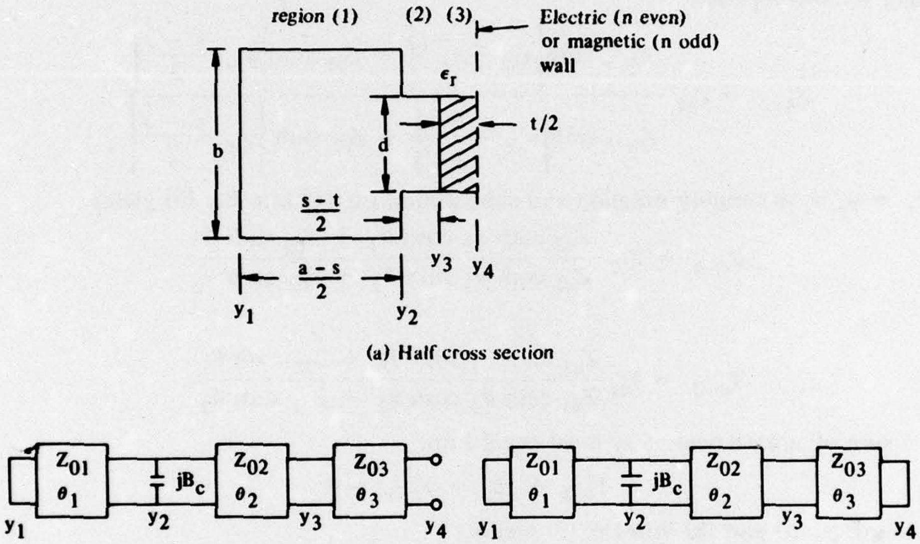


Fig. 2 — Cross section of dielectric slab center loaded, double-ridged waveguide

Cohn's article on ridged waveguide [2] points out that for the homogeneous case (i.e. ϵ_r = 1) the cross section may be treated at the cut-off frequency by assuming that it is an infinitely wide, composite, parallel strip transmission line short-circuited at two points. The resultant electromagnetic field may be considered as an electromagnetic wave traveling from side to side without longitudinal propagation. The resonant conditions can then be solved for the cut-off frequencies of the different TE_{n0} modes.

A similar argument holds for the inhomogeneous case. In addition, the longitudinal propagation constant may be treated as the unknown quantity, and solutions at any frequency may be obtained by separating the wave vector in each region into its transverse and longitudinal components. Since the waveguide configuration is symmetrical, the resonance condition for the transverse wave component will result in an infinite (zero) impedance at the center for n odd (even). Half of a cross section is shown in Fig. 3a, and the equivalent transmission line circuits

for the transverse wave are shown in Fig. 3b for n odd and Fig. 3c for n even. Since the equivalent circuit is a composite, dissipationless, passive line matched at both ends, it is matched at all points. Therefore, the sum of the admittances at the plane y_2 of the effective lumped capacitance due to the ridge wall must equal zero. Within each region, where the regions are shown in Fig. 3a, Z_{0i} is the characteristic impedance, $Y_{0i} = 1/Z_{0i}$ is the characteristic admittance, and θ_i is the transverse electrical length; θ_i is equal to the product of the physical transverse dimension of the region and γ_{yi}, the complex transverse propagation constant. Since all regions are lossless, γ_{yi}, and therefore θ_i, will be purely real or purely imaginary.



(a) Half cross section

(b) Equivalent transmission line circuit for n odd

(c) Equivalent transmission line circuit for n even

Fig. 3 — Half waveguide cross section and equivalent transmission line circuits for transverse wave

The reflected impedance Z presented by a load impedance Z_L terminating a transmission line of characteristic impedance Z_0 with propagation constant γ and length w is [7]

$$Z = Z_0 \frac{(Z_L + Z_0)e^{\gamma w} + (Z_L - Z_0)e^{-\gamma w}}{(Z_L + Z_0)e^{\gamma w} - (Z_L - Z_0)e^{-\gamma w}}. \tag{1}$$

The short circuit at y_1 in Fig. 3b will be reflected back to y_2 as Z_{1-2} where

$$Z_{1-2} = Z_{01} \tanh\left[\gamma_{y1} \frac{a-s}{2}\right] \tag{2}$$

or

$$Y_{1-2} = \gamma_{01} \coth\left[Y_{y1} \frac{a-s}{2}\right]. \tag{3}$$

The open circuit at y_4 will reflect back to y_3 as Z_{4-3} with

$$Z_{4-3} = Z_{03} \coth \left[\gamma_{y3} \frac{t}{2} \right]. \tag{4}$$

Equation (1) can be rewritten in the form

$$Z = Z_0 \frac{Z_L \cosh \gamma w + Z_0 \sinh \gamma w}{Z_L \sinh \gamma w + Z_0 \cosh \gamma w}. \tag{5}$$

Since Z_{4-3} terminates region 2,

$$Z_{4-2} = Z_{02} \frac{Z_{4-3} \cosh \left[\gamma_{y2} \frac{s-t}{2} \right] + Z_{02} \sinh \left[\gamma_{y2} \frac{s-t}{2} \right]}{Z_{4-3} \sinh \left[\gamma_{y2} \frac{s-t}{2} \right] + Z_{02} \sinh \left[\gamma_{y2} \frac{s-t}{2} \right]}. \tag{6}$$

Using $\theta_i = \gamma_{yi} w_i$ to simplify notation and substituting Eq. (4) into Eq. (6) yields

$$Z_{4-2} = Z_{02} \frac{Z_{03} \coth \theta_3 \cosh \theta_2 + Z_{02} \sinh \theta_2}{Z_{03} \coth \theta_3 \sinh \theta_2 + Z_{02} \cosh \theta_2} \tag{7}$$

or

$$Y_{4-2} = Y_{02} \frac{Z_{03} \coth \theta_3 \sinh \theta_2 + Z_{02} \cosh \theta_2}{Z_{03} \coth \theta_3 \cosh \theta_2 + Z_{02} \sinh \theta_2}. \tag{8}$$

Since the sum of admittances at y_2 must equal zero,

$$Y_{1-2} + jB_c + Y_{4-2} = 0. \tag{9}$$

Substituting Eqs. (3) and (8) into Eq. (9) yields

$$Y_{01} \coth \theta_1 + jB_c + Y_{02} \frac{Z_{03} \coth \theta_3 \sinh \theta_2 + Z_{02} \cosh \theta_2}{Z_{03} \coth \theta_3 \cosh \theta_2 + Z_{02} \sinh \theta_2} = 0 \tag{10}$$

or

$$\coth \theta_1 + j \frac{B_c}{Y_{01}} + \frac{Y_{02}}{Y_{01}} \frac{\coth \theta_3 \sinh \theta_2 + \dfrac{Z_{02}}{Z_{03}} \cosh \theta_2}{\coth \theta_3 \cosh \theta_2 + \dfrac{Z_{02}}{Z_{03}} \sinh \theta_2} = 0. \tag{11}$$

Since region 1 and region 2 have the same propagation constant, $\gamma_{y1} = \gamma_{y2}$, the impedances are proportional to the heights:

$$\frac{Z_{02}}{Z_{01}} = \frac{Y_{01}}{Y_{02}} = \frac{d}{b}. \tag{12}$$

Regions 2 and 3 have equal heights, and since the transverse wave is *TE*, the impedance ratio is

$$\frac{Z_{02}}{Z_{03}} = \frac{\gamma_{y3}}{\gamma_{y2}}. \tag{13}$$

The left side of Eq. (11) may be rewritten as a single fraction. All terms in the denominator are finite, so the numerator may be equated to zero. The resultant expression is

$$\left[\frac{b}{d}\sinh\theta_1\right][\gamma_{y2}\cosh\theta_3\sinh\theta_2+\gamma_{y3}\sinh\theta_3\cosh\theta_2]+\left[\cosh\theta_1+j\frac{B_c}{Y_{01}}\sinh\theta_1\right]$$
$$\times[\gamma_{y2}\cosh\theta_3\cosh\theta_2+\gamma_{y3}\sinh\theta_3\sinh\theta_2]=0. \tag{14}$$

Within each region,

$$\gamma_{xi}^2+\gamma_{yi}^2+\gamma_{zi}^2=-\omega^2\mu_0\epsilon_i \tag{15}$$

where

$$\begin{aligned}\epsilon_i &= \epsilon_0 \text{ for } i=1,2\\ &= \epsilon_r\epsilon_0 \text{ for } i=3.\end{aligned}$$

For *TE* modes, $\gamma_{xi}=0$ for all regions and $\gamma_{zi}=j\beta$ for all regions; β is the longitudinal propagation constant (above cutoff) for the waveguide configuration. Substituting

$$\begin{aligned}\gamma_{yi} &= \sqrt{\beta^2-\omega^2\mu_0\epsilon_i} \quad \text{for} \quad \omega^2\mu_0\epsilon_i<\beta^2\\ &= j\sqrt{\omega^2\mu_0\epsilon_i-\beta^2} \quad \text{for} \quad \omega^2\mu_0\epsilon_i\geqslant\beta^2\end{aligned} \tag{16}$$

and

$$\theta_i=\gamma_{yi}w_i$$

with

$$\begin{aligned}w_1 &= 1/2\,(a-s)\\ w_2 &= 1/2\,(s-t)\\ w_3 &= 1/2\,t\end{aligned} \tag{17}$$

into Eq. (14) yields the transcendental equation in β that must be solved for TE_{n0} (n odd) modes. The smallest root of Eq. (14) is the TE_{10} solution, the next root the TE_{30} solution, etc.

For TE_{n0} (n even) modes, the analysis starts with the equivalent transmission line circuit of Fig. 3b and proceeds in a manner similar to the case for n odd. The resultant transcendental equation is

$$\left[\frac{b}{d}\sinh\theta_1\right][\gamma_{y2}\sinh\theta_3\sinh\theta_2+\gamma_{y3}\cosh\theta_3\cosh\theta_2]+\left[\cosh\theta_1+j\frac{B_c}{Y_{01}}\sinh\theta_1\right]$$
$$\times[\gamma_{y2}\sinh\theta_3\cosh\theta_2+\gamma_{y3}\cosh\theta_3\sinh\theta_2]=0 \tag{18}$$

with Eqs. (16) and (17) being applicable.

If $\epsilon_r=1$, it is straightforward to show that Eqs. (14) and (18) reduce to the expressions for the odd and even mode cutoff frequencies, respectively, for double-ridged waveguide [1,2]. Also, if $b=d$, B_c equals zero and Eqs. (14) and (18) result in expressions for the odd- and even-mode propagation constants of dielectric slab center loaded rectangular waveguide identical to those obtained by use of ABCD matrices [6].

The discontinuity-susceptance term B_c/Y_{01} is obtained from the *Waveguide Handbook* [8]. Appendix A gives the necessary equations for calculating B_c/Y_{01} in terms of the waveguide dimensions (from Fig. 2) and the effective wavelength λ_g. Note that λ_g is the wavelength of the wave component which is incident normal to the height change. Therefore λ_g of Appendix A is the wavelength of the transverse wave in regions 1 and 2, namely λ_{y1}.

For standard (i.e. air filled) double-ridged waveguide $\epsilon_r = 1$; thus

$$\gamma_{y1} = \gamma_{y2} = \gamma_{y3} = j\beta_y$$

and $\lambda_{yi} = 2\pi/\beta_y$ is a real constant for a given configuration. However, for the general case $\epsilon_r > 1$ and

$$\gamma_{y1} = \gamma_{y2} \neq \gamma_{y3},$$

with the result that the values of γ_{yi} that satisfy Eq. (14) or (18), subject to (16) and (17), are no longer constant but depend on the frequency. This is of course to be expected, since for any nonhomogeneous waveguide a $1/\sqrt{1 - (f_{c/f})^2}$ term no longer describes the dispersive nature. However, there is another problem because of the inhomogeneity. At all frequencies above cutoff, the transverse wave propagation constant in the dielectric region will be entirely imaginary; i.e.,

$$\gamma_{y3} = j\beta_{y3} \text{ for } \omega > \omega_c.$$

However, there is a critical frequency ω_{crit} (ω_{crit} is greater than the cutoff frequency ω_c; how much greater depends upon the degree of dielectric loading) such that for frequencies greater than ω_{crit} the transverse propagation constant in regions 1 and 2 is real, that is,

$$\gamma_{y1} = \gamma_{y2} = \alpha_{y1} \text{ for } \omega > \omega_{crit}.$$

When $\omega > \omega_{crit}$, the transverse "wave" in these regions is no longer a resonant traveling wave but rather the fields are decaying exponentially away from the dielectric region, and the concept of wavelength in the region of the discontinuity is not meaningful. The expression for the B_c/Y_{01} term from Ref. 8 is no longer applicable; indeed, the validity of the equivalent transmission line circuit for the waveguide height change (a shunt susceptance at the junction of two transmission lines of unequal characteristic impedance) is questionable for operation below cutoff. Also, the calculation for B_c/Y_{01} is based on a model which assumes that the waveguide extends to infinity in both directions away from the height discontinuity; in practice, the assumption is valid if additional mismatches are far enough removed from the height discontinuity so that the local fields have decayed to small proportions. These local fields are the evanescent modes of the fringing fields caused by the height discontinuity, and they decay very rapidly.

Future investigation is planned to model an equivelent circuit of the waveguide height change to include operation below as well as above the cut-off frequency, and to include the proximity effects of waveguide walls and dielectric center loading. However, for this report the following two engineering assumptions are made:

1. The B_c/Y_{01} term can be neglected for frequencies below the critical frequency. Since

$$\frac{B_c}{Y_{01}} \rightarrow 0 \; as \; \omega \rightarrow \omega_{crit}^{(+)}$$

and for $\omega < \omega_{crit}$ the fields of the transverse wave are decaying exponentially in the region of the height discontinuity, a small shunt susceptance term will have only a minor effect on the solution for β. Equations (14) and (18) are transcendental equations and must be solved by some algorithm using trial values of β. If a trial value of β yields an imaginary transverse propagation constant in region 1, the B_c/Y_{01} term is calculated with

$$\gamma_{y1} = j\beta_{y1} \text{ and } \lambda_{y1} = \frac{2\pi}{\beta_{y1}}.$$

If the trial value of β results in γ_{y1} real, the B_c/Y_{01} term is neglected, i.e. set equal to zero, in Eqs. (14) and (18).

2. Proximity effects can be neglected in the calculation of B_c/Y_{01}.

Although the validity of these two assumptions may be questioned from a rigorous theoretical aspect, the close agreement between calculated and measured values of β for different configurations (shown in Figs. 4 and 5) indicates that both assumptions result in accuracy sufficient for most practical applications.

A listing of a computer program to solve for the principal (TE_{10}) mode propagation constant of dielectric slab center loaded double-ridged waveguide is given in Appendix B.

All discussions and calculations thus far have assumed a double-ridged waveguide configuration. For the asymmetric or single-ridged waveguide configurations shown in Fig. 6, Eqs. (14) and (18) remain valid; however, the expression for B_c/Y_{01} must have λ_{y1} replaced by 1/2 λ_{y1}.

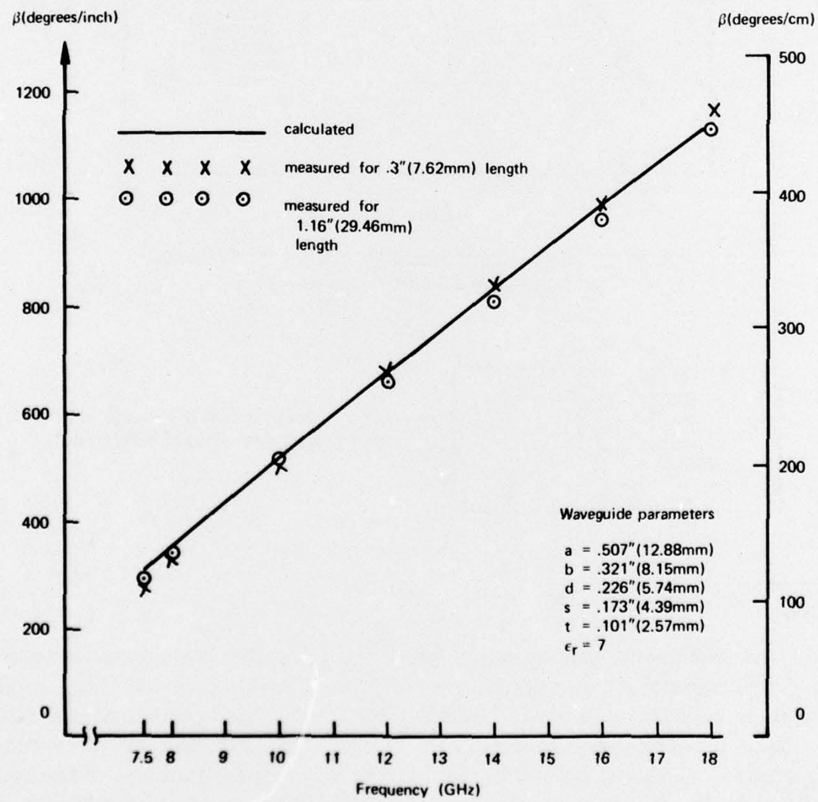


Fig. 4 — Calculated and measured values of β for dielectric slab loaded double-ridged waveguide

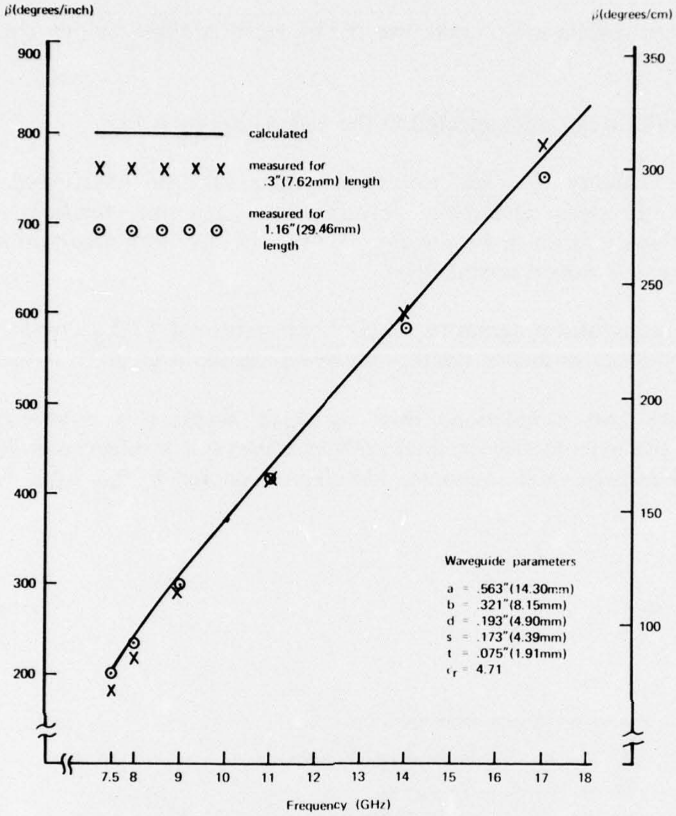


Fig. 5 — Calculated and measured values of β for dielectric slab loaded doubled-ridged waveguide

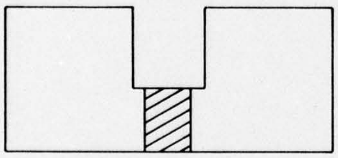


Fig. 6 — Cross section of dielectic slab center loaded single-ridged waveguide

CONCLUSION

Based on the equivalent transmission line circuit for the transverse component of the propagating electromagnetic wave, expressions have been derived for the TE_{n0} mode propagation constants of a dielectric slab center loaded ridged waveguide configuration. These expressions are transcendental equations involving the propagation constant, but they can readily be solved with a computer. Based on the agreement between calculated and measured data, certain assumptions made in the derivation appear valid. The analysis should prove useful in designing transformers to match ridged waveguide to dielectric or ferrite slab center loaded rectangular waveguide.

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Appendix A
EQUIVALENT CIRCUIT FOR A CHANGE IN HEIGHT OF RECTANGULAR WAVEGUIDE

For a height change of rectangular waveguide as shown in Figs. A1a and A1b, the equivalent circuit given by the *Waveguide Handbook** is shown in Fig. A1c. The characteristic admittances of the different height waveguides are Y_0 and Y_0', T is the effective terminal plane, B_c is the effective shunt capacitive susceptance, and λ_g is the wavelength of the propagating wave. The admittance ratio is

$$\frac{Y_0}{Y_0'} = \frac{d}{b} = \alpha$$

and at the terminal plane T

$$\frac{B_c}{Y_0} = \frac{2b}{\lambda_g}\left\{ ln\left(\frac{1-\alpha^2}{4\alpha}\right) + \frac{1}{2}\left(\alpha + \frac{1}{\alpha}\right) ln\left(\frac{1+\alpha}{1-\alpha}\right) + 2\,\frac{A + A' + 2C}{AA' - C^2} + \left(\frac{b}{4\lambda_g}\right)^2 \left(\frac{1-\alpha}{1+\alpha}\right)^{4\alpha} \left(\frac{5\alpha^2 - 1}{1-\alpha^2} + \frac{4}{3}\,\frac{\alpha^2 C}{A}\right)\right\}$$

where

$$A = \left(\frac{1+\alpha}{1-\alpha}\right)^{2\alpha} \frac{1 + \sqrt{1 - \left(\frac{b}{\lambda_g}\right)^2}}{1 - \sqrt{1 - \left(\frac{b}{\lambda_g}\right)^2}} - \frac{1 + 3\alpha^2}{1-\alpha^2}$$

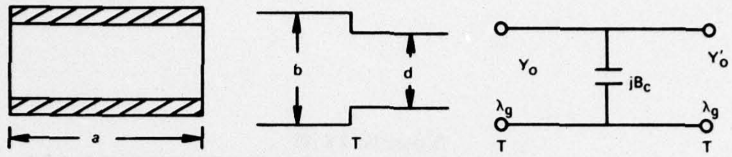
$$A' = \left(\frac{1+\alpha}{1-\alpha}\right)^{2/\alpha} \frac{1 + \sqrt{1 - \left(\frac{d}{\lambda_g}\right)^2}}{1 - \sqrt{1 - \left(\frac{d}{\lambda_g}\right)^2}} + \frac{3 + \alpha^2}{1-\alpha^2}$$

and

$$C = \left(\frac{4\alpha}{1-\alpha^2}\right)^2 .$$

The equivalent circuit is valid for $b/\lambda_g < 1$.

*N. Marcuvitz, *Waveguide Handbook*, MIT Radiation Laboratory Series, McGraw-Hill, New York, 1951.



(a) Cross section (b) Side view (c) Equivalent circuit

Fig. A-1 — Height change of rectangular waveguide and equivalent circuit

Appendix B
FORTRAN LISTING OF COMPUTER PROGRAM

```
C THIS IS PROGRAM DRWGDL.FOR - CWY -OCT 75
        INTEGER RIK
        REAL KXAIR
        PI=3.1415927
        C=2.997925E+08
        R1=39.37008
        R2=2.0*R1
        RRMDI=180.0/(PI*R1)
        C1=(2.0E+09*PI/C)**2
        NEWRUN=0
        TYPE 600
  600   FORMAT (////' PROGRAM DRWGDL/CWY/OCT 75'//' COMPUTES
     1 TE10 CUTOFF FREQUENCIES AND PROPAGATION CONSTANTS OF'/
     2 ' SYMMETRIC DIELECTRIC LOADED DOUBLE RIDGED WAVEGUIDE')
  105   TYPE 605
  605   FORMAT(//' WAVEGUIDE DIMENSIONS IN INCHES - A,B,D,S: '$)
        READ(5,*)A,B,D,S
  106   TYPE 615
  615   FORMAT (/' RELATIVE DIELECTRIC CONSTANT OF CENTER
     1 LOADING = '$)
        READ(5,*)EPSR
        TYPE 625
  625   FORMAT (' WIDTH IN INCHES OF CENTER LOADING = '$)
        ACCEPT 630,T
  630   FORMAT (F8.3)
        IF(T.LT.S)GO TO 108
        TYPE 631
  631   FORMAT (' DIELECTRIC WIDTH MUST BE LESS THAN
     1 RIDGE WIDTH ---- TRY AGAIN')
        GO TO 105
  108   TYPE 606
  606   FORMAT (//' DRWGDL PARAMETERS ------- DIMENSIONS IN
     1 INCHES'/8X' A'9X'B'9X'D'9X'S'12X'T'6X4HEPS'/1X$)
        TYPE 607,A,B,D,S,T,EPSR
  607   FORMAT (4F10.4,F13.4,F10.3)
        R=D/B
        RS=R**2
        IFR=0
        IF(ABS(R-1.0).LT.1.0E-06)IFR=1
        W1=(A-S)/R2
        W2=(S-T)/R2
        W3=T/R2
        CEREST=1.+(1./R-1.)*COS(PI*(A-S)/(2.*A))
        CLREST=CEREST+(EPSR-1.)/R*COS(PI*(A-T)/(2.*A))
        EDCTRY=CLREST/CEREST
        ALCEST=A*CLREST*(R+(1.-R)*SIN(PI*(A-S)/(2.*A)))
C THE ABOVE FOUR QUANTITIES ARE TO BE USED FOR CALCULATING
C APPROXIMATE (STARTING VALUES) OF CUTOFF FREQUENCIES AND
C PROPAGATION CONSTANTS
        IBC=1
```

```
              FREQ=C*R1/(ALCEST*2.0E+09)
              DELBY=0.31*FREQ
              BY=0.0
              GO TO 112
      109     IF(NEWRUN.LT.2)GO TO 210
              IF(FSTART.GT.FCGHZ)GO TO 220
      210         TYPE 635
      635     FORMAT (// FREQUENCIES IN GHZ - START,STOP,INCREMENT: '$)
              READ(5,*)FSTART,FSTOP,DELF
      640     FORMAT (F9.3,1X,F9.3,1X,F9.3)
      220     IF(FSTART.LT.1.0E-13)GO TO 180
              IF(FSTART.GT.FCGHZ)GO TO 230
              TYPE 645
      645     FORMAT(' FREQUENCY MUST BE GREATER THAN CUTOFF')
              GO TO 210
              IF(FSTOP.LT.1.0E-13)FSTOP=FSTART-1.0
      230     TYPE 655
      655     FORMAT (/4X4HFREQ8X4HBETA9X3HGWL7X5HRATIO8X5HKXAIR/
           1 5X3HGHZ6X6HDEG/IN6X6HINCHES4X8HGWL/FSWL7X6HR OR I/)
      110     IFREQ=0
              FREQ=FSTART
      111     IFREQ=IFREQ+1
              BY=PI*2.E+09/C*SQRT(EDCTRY*(FREQ**2+FCGHZ**2))
C THIS IS A FIRST TRY FOR BETA
              DELBY=-0.31*BY
      112     ICROSS=0
              ITAN=0
              IBTRY=0
      115     C1F=C1*FREQ**2
              C1FEP=C1F*EPSR
      120     IBTRY=IBTRY+1
              IF(IBTRY.LT.26)GO TO 122
              TYPE 705
      705     FORMAT (' MORE THAN 25 TRIES AT ROOT')
              GO TO 170
      122     BYSQ=BY**2
              GX3SQ=C1FEP-BYSQ
              GX2SQ=C1F-BYSQ
              GX3=SQRT(ABS(GX3SQ))
              GX2=SQRT(ABS(GX2SQ))
              IF(GX3SQ)130,132,132
      130     CHS3=SINH(GX3*W3)
              CHC3=COSH(GX3*W3)
              IRGX3=1
              GO TO 134
      132     CHS3=SIN(GX3*W3)
              CHC3=COS(GX3*W3)
              IRGX3=-1
      134     CONTINUE
              IF(GX2SQ)136,138,138
```

```
136     CHS2=SINH(GX2*W2)
        CHC2=COSH(GX2*W2)
        CHS1=SINH(GX2*W1)
        CHC1=COSH(GX2*W1)
        KXAIR=GX2
        RIK=1HR
        IRGX2=1
        GO TO 140
138     CHS2=SIN(GX2*W2)
        CHC2=COS(GX2*W2)
        CHS1=SIN(GX2*W1)
        CHC1=COS(GX2*W1)
        IRGX2=-1
        KXAIR=GX2*RRMDI
        RIK=1HI
140     BOY=0.0
        IF(IFR.EQ.1)GO TO 153
        IF(IRGX2.EQ.1)GO TO 153
C CALCULATE B/Y TERM
        P=(1+R)/(1-R)
        GL=2.0*PI/GX2
        P2=SQRT(1.0-(B/(R1*GL))**2)
        P3=SQRT(1.0-(D/(R1*GL))**2)
        PA=P**(2.0*R)*(1.0+P2)/(1.0-P2)-(1.0+3.0*RS)/(1.0-RS)
        PAP=P**(2.0/R)*(1.0+P3)/(1.0-P3)+(3.0+RS)/(1.0-RS)
        PC=((4.0*R)/(1.0-RS))**2
        PT1=ALOG((1.0-RS)/(4.0*R)*P**(0.5*(R+1.0/R)))
        PT2=2.0*(PA+PAP+2.0*PC)/(PA*PAP-PC**2)
        PT3=(B/(R1*4.0*GL))**2*(1.0/P)**(4.0*R)*((5.0*RS
     1 -1.0)/(1.0-RS)+4.0*RS*PC/(3.0*PA))**2
        BOY=2.0*B*(PT1+PT2+PT3)/(R1*GL)
C CALCULATE F(BETA)
153     FBETA=R*(-BOY*CHS1+CHC1)*(GX2/GX3*CHC3*CHC2
     1 +IRGX3*CHS3*CHS2)+IRGX2*CHS1*(GX2/GX3*CHC3*CHS2
     2 +IRGX2*IRGX3*CHS3*CHC2)
        IF(IBC.EQ.1)BY=FREQ
C ROOT SEARCH ROUTINE
        IF(ABS(FBETA).LT.1.0E-08)GO TO 170
        IF(IBTRY.EQ.1)GO TO 163
        IF(ITAN.EQ.1)GO TO 164
        IF(FBETA*FBOLD.LT.0.0)GO TO 161
        IF(ABS(FBETA).GT.ABS(FBOLD))DELBY=-DELBY
        GO TO 162
161     DELBY=-DELBY
        ICROSS=1
162     IF(ICROSS.EQ.1)DELBY=0.5*DELBY
163     BYNEW=BY+DELBY
        IF(ABS((BY-BYNEW)/BY).LT.0.1)ITAN=1
        GO TO 166
164     IF(ABS(BY-BYOLD).LT.1.E-05.AND.FBETA.LT.1.E-06)GO TO 170
```

```
          BYNEW=BY-FBETA*(BY-BYOLD)/(FBETA-FBOLD)
  166     BYOLD=BY
          BY=BYNEW
          FBOLD=FBETA
          IF(IBC.EQ.2)GO TO 120
          FREQ=BY
          BY=0.0
          GO TO 115
  170     IF(IBC.EQ.2)GO TO 175
          FCGHZ=BY
          TYPE 658,FCGHZ,BOY
  658     FORMAT(// TE10 MODE CUTOFF FREQUENCY IN GHZ = 'F7.4'
      1      B/Y = 'F7.3)
          IBC=2
          GO TO 109
  173     CONTINUE
  175     BYDI=BY*RRMDI
          GWL=360.0/BYDI
          FSWL=R1*C/(FREQ*1.0E+09)
          RGLFS=GWL/FSWL
  177     TYPE 660,FREQ,BYDI,GWL,RGLFS,KXAIR,RIK
  660     FORMAT (1X,F7.3,3X,F9.2,3X,F9.4,4X,F8.4,3X,F8.2,1X,A1)
          IF(FREQ.GE.FSTOP)GO TO 180
          FREQ=FREQ+DELF
          GO TO 111
  180     TYPE 665
  665     FORMAT (/// WISH NEW PARAMETERS? NONE=0, ALL=1,
      1 CENTER LOADING=2, FREQ=3      '$)
          ACCEPT 670,NEWRUN
  670     FORMAT (I1)
          GO TO(199,105,106,210,180)NEWRUN+1
  199     CONTINUE
          END
```